

## Research Article

# Effects of Channel Estimation on Multiuser Virtual MIMO-OFDMA Relay-Based Networks

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A practical multiuser cooperative transmission scheme denoted as *Virtual Maximum Ratio Transmission* (VMRT) for multiple-input multiple-output-orthogonal frequency division multiple access (MIMO-OFDMA) relay-based networks is proposed and evaluated in the presence of a realistic channel estimation algorithm and using low-density parity-check (LDPC) codes. It is shown that this scheme is robust against channel estimation errors. It offers diversity and array gain, keeping the complexity low with a multiuser and multiantenna channel estimation algorithm that is simple and efficient. In addition, the combination with LDPC codes provides improved gains; diversity gains larger than 6 dB can be easily obtained with a reduced number of relays. Thus, this scheme can be used to extend coverage or increase system throughput by using simple cooperative OFDMA-based relays.

## 1. Introduction

The idea of increasing reliability, coverage, and/or capacity in future wireless networks by using cooperative single-antenna relays to reach users' terminals has recently attracted much attention [1–15]. In addition, Multiple-Input Multiple-Output (MIMO) technology has demonstrated that it is a good approach to increase capacity [16, 17]; together with Orthogonal Frequency Division Multiplexing (OFDM) [18] or Orthogonal Frequency Division Multiple Access (OFDMA) [19], MIMO techniques can also provide increased reliability. The right combination of all these elements would lead to a considerable improvement of system performance.

Relay schemes can be categorized into three different groups: *Amplify-and-Forward* (AF) [3, 4, 8, 10–13], *Compress-and-Forward* (CF) [5, 20], and *Decode-and-Forward* (DF) [1–3, 6, 7, 9, 15]. In the AF schemes, relays amplify (and maybe transform [4]) the received signal and broadcast it to the destination. These schemes can be appropriate to extend coverage or to solve the problem of attenuation faced by receivers. Furthermore, some spatial diversity can be

provided [1, 6]. In the CF, the relay transmits a quantized and compressed version of the received signal to the destination, and the destination decodes the signal by combining it with its own received signal. These schemes can exploit the redundancy between source and destination, and they assume that the source is able to reach the destination. In the last group, relays in the DF strategy decode the received signal and re-encode (and possibly transform/adapt) the information and send it to destinations. In [5], it is shown that CF strategies outperform DF when the relays are closer to the destination, and DF obtains larger throughput when relays are closer to the source. Since among the applications of our scheme is coverage extension (which imposes that the source cannot directly reach the destinations) and the use of simple relays, in this paper, we adopt this last strategy because in these scenarios better performance can be achieved by DF.

In [8], it is shown that the conventional Maximum Ratio Combining (MRC) is the optimum detection scheme for the AF strategy and also that it can achieve full diversity order of  $K + 1$ , where  $K$  is the number of relays, whereas for the DF strategy, the optimum is the Maximum Likelihood (ML) detector [1, 9]. As recognized in [1], performance analysis

and implementation of said detector are quite complicated and thus a suboptimum combiner termed as  $\lambda$ -MRC was derived. Another suboptimum detector is the cooperative MRC (C-MRC) [10] and link adaptive regeneration (LAR) [11]. In these works, collaboration is performed at the destination, namely, the receiver treats the relays as a multiple-source transmitter and combines the multiple received signal adequately to obtain the best performance. If we also take relays into account in the design, we can improve the throughput and lower the outage probability by selecting the best relays to transmit from [12, 13] (for the AF strategy) and [7] (for the DF). Going further, we can consider the relays as a virtual multiple-input transmitter (if cooperation is used), and thus leverage on it to improve destination (user) performance. In [14, 15], the relays are used as a beamformer where full or partial channel state information (CSI) is needed on all the elements, and a joint optimization is performed to obtain the best results at the destination. However, in a practical scenario, knowledge of CSI (even partial) from all the network elements at the source (CSI-T) is not possible, and moreover, it needs to be estimated and errors might occur.

In addition, the time-frequency structure of OFDMA offers flexibility in terms of multiuser resource management and advantages in terms of dealing with multipath wireless channel effects. Moreover, next generation wireless mobile networks will use some combination of the OFDMA transmission technique [21]. For this reason, in this paper OFDMA has been selected in combination with MIMO to offer a global system design with high data rate capacity and flexibility in terms of accommodating multiple users.

On the other hand, channel-coding schemes are able to drastically improve performance, while channel estimation errors may seriously affect them. Although capacity-approaching codes such as the low-density parity-check (LDPC) were proposed long ago [22], these codes have recently attracted much attention due to their efficient implementations [23] and large coding gains [24].

In [25], the authors propose and analyze a practical transmission scheme with the DF strategy taking the relays as a *Virtual Multiple-Input Transceiver* (VMIT). However, perfect and instantaneous CSI is assumed and no channel code is used. In this paper, we design and examine the performance of this scheme in the presence of a realistic and practical channel estimation algorithm and with the use of powerful LDPC codes. The acquisition of channel state information in a multiuser VMIT must be carried out in an efficient and simple way in order not to have a serious impact on bandwidth efficiency. Lowering the pilot overhead and the complexity of the channel estimation scheme adopted in all the receivers in the system is of paramount importance, and as the number of users and relays increases, it becomes mandatory. Thus, the proposal in [26] is used to fit requirements.

Our contributions in this paper are

- (i) the comparison of different practical transmission schemes in a MIMO-OFDMA-relay-based network with a base station with  $N_t$  transmit antennas, using

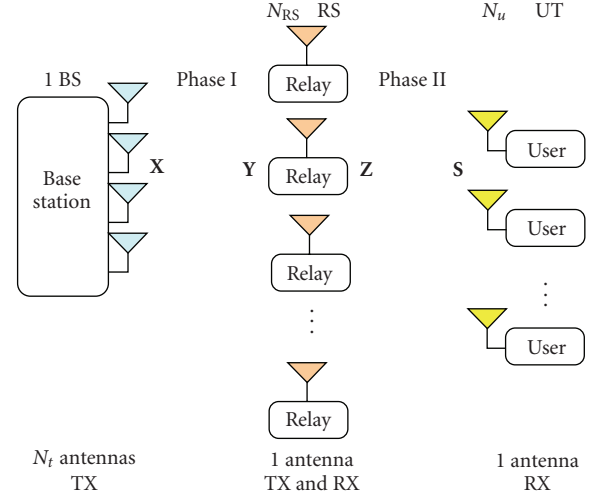


FIGURE 1: Scenario used in the paper.

the *Decode-and-Forward* strategy, and LDPC channel codes and keeping the complexity low;

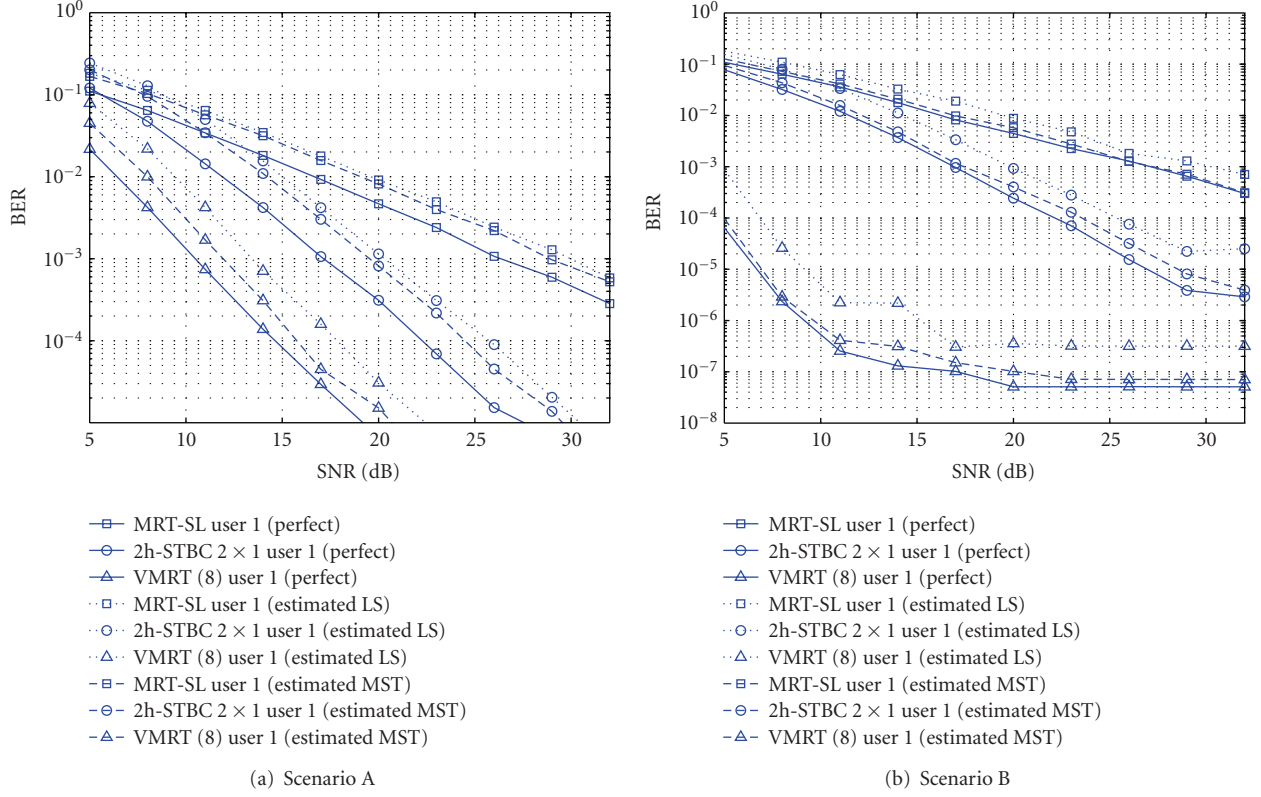
- (ii) a proposal for the transmission over this network that obtains diversity and array gain at the users' terminals with increase in system performance and reliability with no CSI-T either at the base station or at the relays and with low complexity;
- (iii) the evaluation of these schemes when there is degradation in the CSI due to the use of a realistic channel estimation algorithm;
- (iv) the evaluation of the LDPC codes in such two-hop distributed systems.

The remainder of this paper is organized as follows. First, in Section 2, a description of the scenario and the system model is presented. Next, in Sections 3 and 4, the proposed scheme and the proposed channel estimation are described and summarized, respectively. In Section 5, the results are presented and discussed. Finally, some conclusions are drawn in Section 6.

**Notations.** Throughout the paper the following notation will be used. Bold capitals and bold face for matrices and vectors, respectively.  $E_y\{x\}$  denotes expectation of  $x$  over  $y$  and  $|\mathbf{h}|$  and  $\|\mathbf{h}\|$  account for the absolute value and the square of the 2-norm of  $\mathbf{h}$ , respectively. The square of this norm will be denoted in the paper as gain ( $\mathbf{h}^H\mathbf{h}$ ).  $\mathbf{I}_N$  is the identity matrix of size  $N$ , and  $\text{diag}\{\mathbf{x}\}$  is a diagonal matrix containing  $\mathbf{x}$  in its diagonal and 0 elsewhere.

## 2. Description of the Scenario and System Model

The reference scenario is shown in Figure 1 and is based on a base station (BS) with  $N_t$  transmit antennas,  $N_{RS}$  cooperative relay stations (RSs), each one with only one antenna for transmission and reception, and  $N_u$  user's terminals (UT), also with one receive antenna each. We assume that the users cannot be reached by the BS directly. The strategy used

FIGURE 2: Effect of channel estimation: uncoded QPSK, SUI-3 channel;  $N_t = 4$ ,  $N_{RS} = 8$ ,  $N_{VMRT} = 8$ ,  $N_u = 4$ .

is the *Decode-and-Forward* in a half-duplex transmission; that is, in phase I, the BS transmits and RSs receive first link/hop and, in phase II, the relays transmit and UTs receive second link/hop. The system uses  $N$  subcarriers that can be allocated to different users in an OFDMA transmission; that is, different UTs use disjoint sets of  $N_i$  orthogonal subcarriers. We assume, for simplicity and without loss of generality, that the subcarriers used in the link BS-RS are the same as in the link RS-UT. The algorithm or policy for the scheduler to assign subcarriers is beyond the scope of the paper. We will consider the transmission of  $N_s$  OFDMA symbols as a block and denote a packet as a group of several blocks. In general,  $N_s$  can take any value. However, for the space-time block code-(STBC-) based schemes that we are proposing, the block size must necessarily equal the number of transmit antennas, that is,  $N_s = N_t$ . This is because we are proposing the use of full-rate STBC.

The frequency-domain transmitted signal from the BS is

$$\mathbf{X}^k = \mathbf{V}\mathbf{C}^k, \quad (1)$$

where  $\mathbf{X}^k \in \mathbb{C}^{N_t \times N_s}$  is the signal transmitted from the  $N_t$  antennas at  $k$ th subcarrier during block of  $N_s$  OFDMA symbols,  $\mathbf{V} \in \mathbb{C}^{N_t \times N_s}$  is a generic precoding matrix  $k$ , and  $\mathbf{C}^k \in \mathbb{C}^{N_s \times N_s}$  are the complex base band data to be sent on the  $k$ th subcarrier by all the transmit antennas, assumed here to be  $M$ -QAM or  $M$ -PSK modulated without loss of generality.

Next, the frequency-domain received signal at the  $i$ th relay on the  $k$ th subcarrier after discrete fourier transform (DFT) and discarding the cyclic prefix (CP) can be written as

$$\mathbf{y}_i^k = \mathbf{h}_i^k \mathbf{X}^k + \boldsymbol{\psi}^k, \quad (2)$$

where  $\mathbf{y}_i^k \in \mathbb{C}^{1 \times N_s}$  is the received signal by relay  $i$  at subcarrier  $k$ ,  $\mathbf{h}_i^k \in \mathbb{C}^{1 \times N_t}$  is the channel frequency response for relay  $i$  at subcarrier  $k$  from all the transmit antennas ( $N_t$ ), and  $\boldsymbol{\psi}^k$  is the zero-mean additive white Gaussian noise (AWGN) vector, with each component ( $k$ ) with variance  $\sigma_i^2$ . We can arrange the signal received by all the relays in a matrix form as

$$\mathbf{Y}^k = \mathbf{H}^k \mathbf{X}^k + \boldsymbol{\Psi}^k, \quad (3)$$

where  $\mathbf{Y}^k \in \mathbb{C}^{N_{RS} \times N_s}$  is the received signal by all the relays at  $k$ th subcarrier during a block of  $N_s$  OFDMA symbols, the matrix  $\mathbf{H}^k \in \mathbb{C}^{N_{RS} \times N_t} = [\mathbf{h}_1^k; \mathbf{h}_2^k; \dots; \mathbf{h}_{N_{RS}}^k]$  accounts for the channel frequency response on  $k$ th subcarrier, and  $\boldsymbol{\Psi}^k \in \mathbb{C}^{N_{RS} \times N_s}$  contains the zero-mean AWGN. The  $k$ th subcarrier can be assigned to any user by the scheduler.

For the second hop, namely, from RS to UT, the frequency-domain joint transmitted signal (It should be noted that each relay transmits one of the rows of the joint matrix  $\mathbf{Z}^k$ . Thus, the precoding matrix  $\mathbf{W}$  must be diagonal, otherwise relays would have to share transmission information, and therefore the complexity would increase,

which is not the case) is

$$\mathbf{Z}^k = \mathbf{W}\tilde{\mathbf{X}}^k, \quad (4)$$

where  $\mathbf{Z}^k \in \mathbb{C}^{N_{RS} \times N_s}$  is the signal transmitted by relays at  $k$ th subcarrier during the block of  $N_s$  OFDMA symbols,  $\mathbf{W} \in \mathbb{C}^{N_{RS} \times N_{RS}}$  is a new generic precoding matrix for the second hop, and  $\tilde{\mathbf{X}}^k$  is the estimated  $\mathbf{X}^k$  from received  $\mathbf{Y}^k$  and the remodulated transmitted signal. Since the relays are equipped with only one antenna, the estimated signal is performed in a multiple-input single-output (MISO) way by each relay. In this paper, a simple zero-forcing (ZF) equalization and detection is used for reducing complexity at relays and user's terminals. This yields the following frequency-domain received signal at user's terminal  $u$

$$\mathbf{s}_u^k = \mathbf{h}_u^k \mathbf{Z}^k + \phi_u^k, \quad (5)$$

where  $\mathbf{s}_u^k \in \mathbb{C}^{1 \times N_s}$  is the received signal for user  $u$  at  $k$ th subcarrier during the block of  $N_s$  OFDM symbols,  $\mathbf{h}_u^k \in \mathbb{C}^{1 \times N_{RS}}$  is the channel frequency response for user  $u$  from the  $N_{RS}$  relays at  $k$ th subcarrier, and  $\phi_u^k \in \mathbb{C}^{1 \times N_s}$  is a second AWGN noise vector for subcarrier  $k$  with each component of variance  $\sigma_u'^2$ . Again, grouping all the received signals by users into a matrix yields

$$\mathbf{S}^k = \mathcal{H}^k \mathbf{Z}^k + \Phi^k, \quad (6)$$

$\mathbf{S}^k \in \mathbb{C}^{N_u \times N_s}$  being the received signal by all the users on subcarrier  $k$  during the block of  $N_s$ , the matrix  $\mathcal{H}^k \in \mathbb{C}^{N_u \times N_{RS}}$  the channel frequency response from relays to users at  $k$ th subcarrier, and  $\Phi^k \in \mathbb{C}^{N_u \times N_s}$  a second AWGN matrix. Note that since the system uses OFDMA, at reception, each UT selects the subcarriers with data allocated to it among all the received subcarriers.

In this paper, the evaluation of the performance is based on the bit error rate (BER) as a measurement over different Signal-to-noise ratios (SNR). In the scenarios, there are two different links, one from BS to RS and another from RS to UT. Thus, we define the SNR for each link separately. In addition, since the system is MIMO-OFDMA-based, there

will exist  $N_t$  different channels (in the first link) over  $N$  different subcarriers. For these reasons, the average SNR per link is defined as

$$\overline{\text{SNR}} = E_k \left\{ E_i \left[ \frac{|\mathbf{X}_i^k|^2}{\sigma_i^2} \right] \right\}, \quad \begin{matrix} k = 0 \dots N-1, \\ i = 0 \dots N_t-1. \end{matrix} \quad (7)$$

Looking at (7), the SNR is calculated, averaging the signal  $\mathbf{X}_i^k$  over the transmit antennas and the subcarriers. In this way, a single value per link is obtained to associate with the performance in a given scenario. When transmitting from relays, we will have  $N_{RS}$  different channels, and in (7),  $N_t$  should be replaced by the number of transmitting relays for the scheme ( $N_{RS}$ ) and  $\sigma$  by  $\sigma'$ .

It should be noted here that the SNR is used as a way of describing different scenarios for evaluation purposes, but it is not a parameter that needs to be estimated to perform the transmission.

**2.1. A Non-CSI-T Scheme: 2-Hop Space-Time Block Code (2h-STBC).** Although Virtual Maximum Ratio Transmission (VMRT) does not need CSI-T at the relays because the UTs compute the beamforming weights (see Section 3), the selected terminal (and only the selected one) must send its weights to the relays regularly. For this reason, in order to compare and evaluate the impact of channel estimation errors and the use of LDPC codes of the proposed VMRT with the case where no CSI-T is needed, a 2-hop space-time block code is used, denoted as *2h-STBC* throughout the paper; this encoding scheme uses STBC codes in both links. In phase I the BS transmits using Alamouti [27] when using 2 antennas or when using 4 or 8 antennas, [28, 29] which is denoted as “*Alamoutitation*” in [29]. For this scheme, the precoding matrix in (1) is  $\mathbf{V} = \mathbf{I}_{N_t}$  and the number of OFDMA symbols per block ( $N_s$ ) is set to  $N_t$ . Thus, the transmitted signal can be written as

$$\mathbf{X}^k \big|_{\text{STBC}} = \mathbf{C}_\alpha^k, \quad (8)$$

with  $\alpha = 2, 4, 8$  when  $N_s = 2, 4$ , or 8, respectively, and

$$\begin{aligned} \mathbf{C}_2^k &= \begin{bmatrix} c^k(1) & -c^k(2)^* \\ c^k(2) & c^k(1)^* \end{bmatrix}, \\ \mathbf{C}_4^k &= \begin{bmatrix} c^k(1) & c^k(2)^* & c^k(3)^* & c^k(4) \\ c^k(2) & -c^k(1)^* & c^k(4)^* & -c^k(3) \\ c^k(3) & c^k(4)^* & -c^k(1)^* & -c^k(2) \\ c^k(4) & -c^k(3)^* & -c^k(2)^* & c^k(1) \end{bmatrix}, \\ \mathbf{C}_8^k &= \begin{bmatrix} c^k(1) & c^k(2)^* & c^k(3)^* & c^k(4) & c^k(5)^* & c^k(6) & c^k(7) & c^k(8)^* \\ c^k(2) & -c^k(1)^* & c^k(4)^* & -c^k(3) & c^k(6)^* & -c^k(5) & c^k(8) & -c^k(7)^* \\ c^k(3) & c^k(4)^* & -c^k(1)^* & -c^k(2) & c^k(7)^* & c^k(8) & -c^k(5) & -c^k(6)^* \\ c^k(4) & -c^k(3)^* & -c^k(2)^* & c^k(1) & c^k(8)^* & -c^k(7) & -c^k(6) & -c^k(5)^* \\ c^k(5) & c^k(6)^* & c^k(7)^* & c^k(8) & -c^k(1)^* & -c^k(2) & -c^k(3) & -c^k(4)^* \\ c^k(6) & c^k(5)^* & c^k(8)^* & -c^k(7) & -c^k(2)^* & c^k(1) & -c^k(4) & c^k(3) \\ c^k(7) & c^k(8)^* & -c^k(5)^* & -c^k(6) & -c^k(3)^* & -c^k(4) & c^k(1) & -c^k(2)^* \\ c^k(8) & -c^k(7)^* & -c^k(6)^* & c^k(5) & -c^k(4)^* & c^k(3) & c^k(2) & -c^k(1)^* \end{bmatrix}, \end{aligned} \quad (9)$$

being the matrices containing the data to be sent.  $c^k(n)$  are the data on subcarrier  $k$  at OFDMA symbol  $n$  ( $n = 1 \cdots N_s$ ).

All the relays will receive the signal, and thus they are able to decode it, *that is*,  $\mathbf{y}_i^k$  in (2). Grouping all the received signals by all relays, (3) yields

$$\mathbf{Y}^k \Big|_{\text{STBC}} = \mathbf{H}^k \mathbf{X}^k \Big|_{\text{STBC}} + \mathbf{\Psi}^k. \quad (10)$$

Therefore, a cooperative *Virtual STBC* transmission can be carried out from RS in phase II, assuming that the RSs

are numbered and perfectly synchronized. Now, each relay, or a group of  $N_{R2}$  relays, acts as an antenna re-encoding the received signal  $\mathbf{y}_i^k$  into  $\check{\mathbf{x}}_i^k$ . Again, in the general expression of (4), the pre-coding matrix is  $\mathbf{W} = \mathbf{I}_{N_{RS}}$ , and thus arranging all the transmitted signals from the relays into a matrix form, we obtain

$$\mathbf{Z}^k \Big|_{2h\text{-STBC}} = \check{\mathbf{X}}_\beta^k \quad (11)$$

with  $\beta = 2, 4, 8$  for  $N_{R2} = 2, 4$ , or 8, respectively, and

$$\begin{aligned} \check{\mathbf{X}}_2^k &= \begin{bmatrix} \check{x}_1^k(1) & -\check{x}_1^k(2)^* \\ \check{x}_2^k(2) & \check{x}_2^k(1)^* \end{bmatrix}, \\ \check{\mathbf{X}}_4^k &= \begin{bmatrix} \check{x}_1^k(1) & \check{x}_1^k(2)^* & \check{x}_1^k(3)^* & \check{x}_1^k(4) \\ \check{x}_2^k(2) & -\check{x}_2^k(1)^* & \check{x}_2^k(4)^* & -\check{x}_2^k(3) \\ \check{x}_3^k(3) & \check{x}_3^k(4)^* & -\check{x}_3^k(1)^* & -\check{x}_3^k(2) \\ \check{x}_4^k(4) & -\check{x}_4^k(3)^* & -\check{x}_4^k(2)^* & \check{x}_4^k(1) \end{bmatrix}, \\ \check{\mathbf{X}}_8^k &= \begin{bmatrix} \check{x}_1^k(1) & \check{x}_1^k(2)^* & \check{x}_1^k(3)^* & \check{x}_1^k(4) & \check{x}_1^k(5)^* & \check{x}_1^k(6) & \check{x}_1^k(7) & \check{x}_1^k(8)^* \\ \check{x}_2^k(2) & -\check{x}_2^k(1)^* & \check{x}_2^k(4)^* & -\check{x}_2^k(3) & \check{x}_2^k(6)^* & -\check{x}_2^k(5) & \check{x}_2^k(8) & -\check{x}_2^k(7)^* \\ \check{x}_3^k(3) & \check{x}_3^k(4)^* & -\check{x}_3^k(1)^* & -\check{x}_3^k(2) & \check{x}_3^k(7)^* & \check{x}_3^k(8) & -\check{x}_3^k(5) & -\check{x}_3^k(6)^* \\ \check{x}_4^k(4) & -\check{x}_4^k(3)^* & -\check{x}_4^k(2)^* & \check{x}_4^k(1) & \check{x}_4^k(8)^* & -\check{x}_4^k(7) & -\check{x}_4^k(6) & -\check{x}_4^k(5)^* \\ \check{x}_5^k(5) & -\check{x}_5^k(6)^* & \check{x}_5^k(7)^* & \check{x}_5^k(8) & -\check{x}_5^k(1)^* & -\check{x}_5^k(2) & -\check{x}_5^k(3) & -\check{x}_5^k(4)^* \\ \check{x}_6^k(6) & \check{x}_6^k(5)^* & \check{x}_6^k(8)^* & -\check{x}_6^k(7) & -\check{x}_6^k(2)^* & \check{x}_6^k(1) & -\check{x}_6^k(4) & \check{x}_6^k(3)^* \\ \check{x}_7^k(7) & \check{x}_7^k(8)^* & -\check{x}_7^k(5)^* & -\check{x}_7^k(6) & -\check{x}_7^k(3)^* & -\check{x}_7^k(4) & \check{x}_7^k(1) & -\check{x}_7^k(2)^* \\ \check{x}_8^k(8) & -\check{x}_8^k(7)^* & -\check{x}_8^k(6)^* & \check{x}_8^k(5) & -\check{x}_8^k(4)^* & \check{x}_8^k(3) & \check{x}_8^k(2) & -\check{x}_8^k(1)^* \end{bmatrix}, \end{aligned} \quad (12)$$

with  $\check{x}_i^k(n)$  being the re-encoded signal transmitted by the RS  $i$  at  $n$ th OFDMA symbol ( $n = 1 \cdots N_s$ ). Some observations must be pointed out here. The first one is that a different number of transmit elements can be used on each link; that is,  $N_t$  can be different from  $N_{RS}$  and  $N_{R2}$ ; in fact, usually  $N_{RS}, N_{R2} > N_t$ . Since all the relays decode the transmitted signal by BS, the increase in the number of *virtual* transmitters (relays) will exploit diversity and array gains, and the second one is that the transmitted information by relays may not be orthogonal anymore because each relay decodes the received data and some errors can appear. Thus, some degradation in the performance can be expected at the user's end, especially for the channel estimation algorithm and/or LDPC codes. This scheme is the simplest method to obtain diversity from both links, so we will use it as a reference. Moreover, it can be noted that no CSI-T is needed, but rather only channel state information at the receiver (CSI-R) for coherent demodulation, at both links.

### 3. Virtual Maximum Ratio Transmission (VMRT)

In order to obtain diversity in both links with reduced complexity and CSI in all the elements in the network, in

[25], the following scheme is proposed, denoted as *Virtual Maximum Ratio Transmission*, because the relays are used as a cooperative virtual beamformer. In this scheme, the BS uses STBC (2, 4, or 8 scheme) to transmit to relays as in the *2h-STBC* scheme. Therefore, the signal model is the same until the first hop as in *2h-STBC*. In the second hop, instead of using an STBC again, here, the relays are configured as a virtual beamformer, and they conform the signal to the user with the best quality. The beamformer can be performed with all the relays or a group of  $N_{VMRT}$ . In order to reduce the complexity at the relays and the CSI requirements, we use an approach similar to the one of [30]. The step-by-step procedure is as follows.

- (1) Users' terminals estimate the channel matrix and compute the Maximum Ratio Transmission (MRT) weights.
- (2) Each UT computes the link *quality* ( $q_j$ ), only over its subcarriers; that is,  $1/q_j = \max_k \{\text{BER}_j^k\}$ ,  $k \in \mathcal{N}_j$ , where  $\mathcal{N}_j$  is the set of subcarriers allocated to user  $j$  and  $\text{BER}_j^k$  is the estimated BER at subcarrier  $k$  for  $j$ th terminal. (e.g., for QPSK modulation, BER at subcarrier  $k$  for  $i$ th terminal ( $\text{BER}_i^k$ ) can be estimated as  $\text{erfc}(\sqrt{(\sigma_i^2/2)\eta_i^k\eta_i^{kH}}) - (1/4)(\text{erfc}((\sigma_i^2/2)\eta_i^k\eta_i^{kH}))^2$ ,



whereas for 64-QAM, BER can be estimated as  $(1/4) \operatorname{erfc}(\sqrt{(3\sigma_i^2/5)} h_i^k h_i^{kH})$ , where  $\operatorname{erfc}(x) = (2/\sqrt{\pi}) \int_x^\infty e^{-t^2} dt$ .

- (3) UTs broadcast their *quality* to relays; it should be noted that this value is only a scalar per user.
- (4) All RSs receive this value from each UT, and according to the *minimax* BER criterion, the one with the minimum maximum BER is scheduled to transmit. As was shown in [25], this metric is the one which obtains the performance closest to the optimum. If qualities are sorted out in ascending order so that  $q_1 > q_2 > \dots > q_{N_u}$ , the UT with  $q_1$  is selected.
- (5) One RS can act as *coordinator* and informs the selected UT.
- (6) The selected user sends the pre-coding weights vector to relays to obtain the already calculated fed-back quality ( $q_1$ )
- (7) Each RS uses the adequate weight to perform the cooperative Virtual Maximum Ratio Transmission.

Thus, transmitted signal  $\mathbf{Z}^k$  in (4) will use (10) with  $\mathbf{W} = \operatorname{diag}\{\mathbf{w}\}$ , calculated by using the *minimax* BER criterion

$$\mathbf{w} = \frac{h_{j^*}^{k^*}}{\|h_{j^*}^{k^*}\|}, \quad \begin{aligned} j^* &= \arg \min \{ \max_k \{ \operatorname{BER}_j^k \} \}, & k \in \mathcal{N}_j, \\ k^* &= \arg \max_k \{ \operatorname{BER}_j^k \}, & j = 1 \dots N_u. \end{aligned} \quad (13)$$

It should be noted that, although it is a multicarrier system, only one weight per transmit antenna is needed since using the *minimax BER criterion*, the best weight per transmit antenna for all the subcarriers is obtained (Note that  $\mathbf{w}$  is not dependent on the subcarrier index  $k$ ). In this way, the required feedback is reduced and is independent of the number of subcarriers.

Statistically, if the average SNR is the same for all terminals and if the channel is ergodic, then the performance is identical for all users since all of them will sometimes experience the best *quality* channel on the average. By using this scheme, diversity is exploited in both links, especially on the second one, since usually the number of RS is higher than the number of transmit antennas. The reader is referred to [25] for more details.

#### 4. Channel Estimation

The use of coherent demodulation implies the knowledge of the CSI-R at the receivers. The initial proposals for pilot-aided channel estimation schemes for MIMO-OFDM transformed the problem of estimating overlapping channels in the estimation of multiple single-input single-output (SISO) OFDM channels. This was achieved by allocating dedicated pilot subcarriers to each transmit antenna. The receiver estimates each channel from the pilot subcarriers belonging to each transmit antenna, and then it applies an interpolator to get the full channel estimate [31, 32]. This type of pilot allocation can be found in the fixed WIMAX standard [33].

Although this type of pilot allocation simplifies the channel estimation, it presents some drawbacks. As the number of transmit antennas increases, the spectral efficiency decreases considerably since a large number of subcarriers will be assigned exclusively to transmit pilots. Moreover, the fact that the pilot subcarriers are not loaded in any except the transmit antenna for which the subcarrier is allocated increases the critical peak-to-average power ratio (PAPR) parameter [34], which strongly impacts on the performance of the power amplifier.

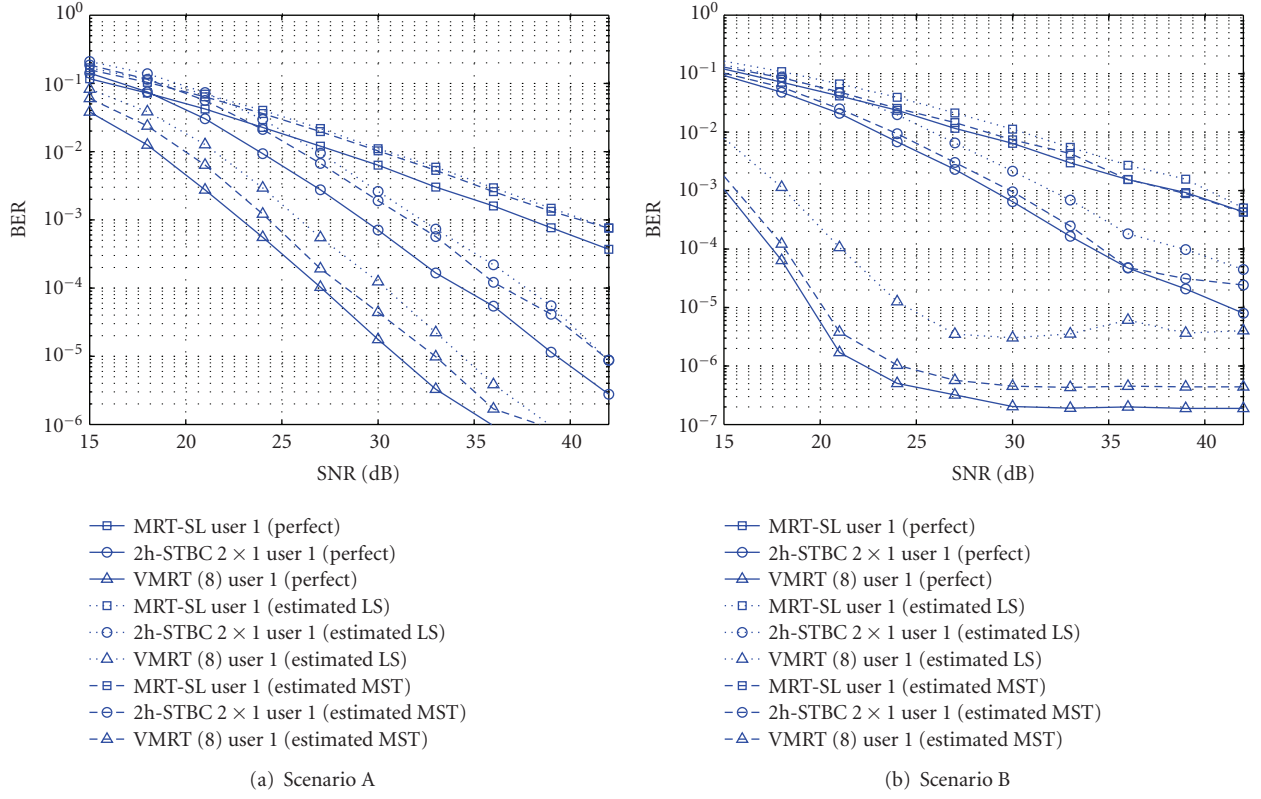
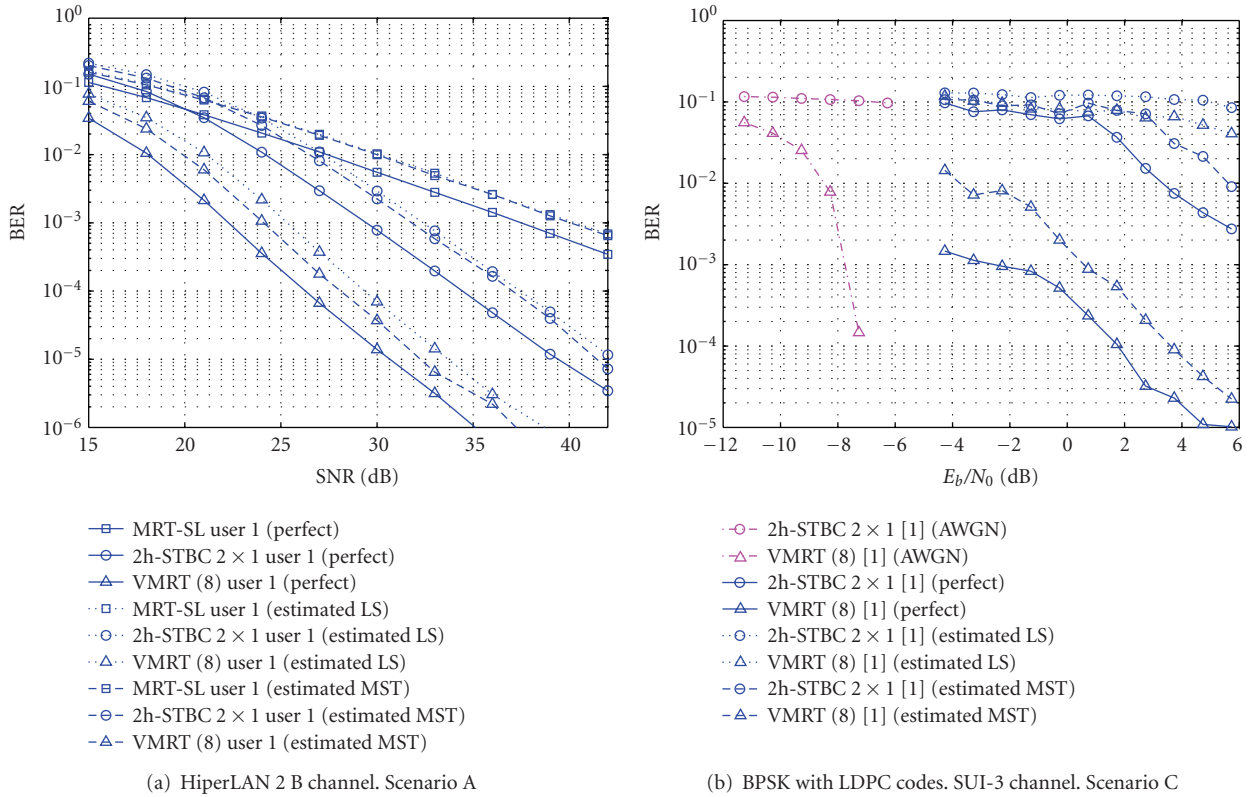
In our scenario, where the BS can be equipped with several antennas or the VMIT can be configured as a large number of transmit antennas ( $N_{\text{VMIT}}$ ), the pilots must be sent efficiently to minimize the decrease in the system's efficiency but still enable the receivers to estimate all the channels accurately, with minimum cochannel interference.

A pilot-aided channel estimation scheme that attempted to minimize the cochannel interference was published in [35]. The proposed algorithm exhibits a high computational load. A simplified and enhanced algorithm, introducing a data-aided scheme for the data transmission mode, is presented in [36]. In [37], overlapped pilots are proposed for channel estimation where different transmitters use the same pilot subcarriers, avoiding the decrease in efficiency with an increasing number of transmitters. However, the performance results are not very favorable. The topic attracted significant attention and has been the focus of research in multiple publications [38–40] and references therein.

The design of training symbols and pilot sequences with the ability to decouple the cochannel interference and minimize the channel estimation mean square error (MSE) for MIMO-OFDM was addressed in several publications [36, 41, 42]. In addition, the use of different orthogonal sequences was addressed in several works. The use of Hadamard sequences was proposed in [34, 43], while the Golay sequences were considered in [44] and complex exponential sequences were investigated in [45, 46]. The time-domain channel estimation schemes have not received much attention due to the insurmountable fact that the equalization is performed in the frequency domain. Nevertheless, some research on the topic can be found in the literature.

The design of the pilot sequences is explored in [47, 48]. The pilot-carrying received symbols are processed to explore the correlation among the several channel impulse response (CIR) replicas to reduce the noise in the estimate. The use of superimposed pseudorandom pilot sequences was investigated in [47, 49]. In these schemes, the CIR estimate is obtained through the correlation of the received symbols with copies of transmitted pseudorandom sequences that are stored in the receiver (known a priori).

Although published work on time-domain channel estimation showed that the estimation process can be performed directly in time domain, due to the common frequency-domain pilot arrangement, most of the publications on the topic of pilot-aided channel estimation use the frequency-domain least squares (LS) estimates as the starting point for the estimation process. The results in [50] show that this

FIGURE 3: Effect of channel estimation: uncoded 64QAM, SUI-3 channel;  $N_t = 4$ ,  $N_{RS} = 8$ ,  $N_{VMRT} = 8$ ,  $N_u = 4$ .FIGURE 4: Performance results: effect of channel estimation;  $N_t = 4$ ,  $N_{RS} = 8$ ,  $N_u = 4$ ,  $N_{VMRT} = 8$ .

operation can be performed in time-domain by a simple linear operation on the received signal.

In this paper, we adopt the MIMO-OFDM pilot sequence design, where the same set of subcarriers conveys pilots for all antennas, and the pilot sequence corresponding to each transmit antenna is coded with different orthogonal phase-shifting sequences. This sequence design is proven to be optimal in [42]. The pilot design, together with the associated channel estimation method [26], succeeds in estimating all the channels involved in the transmission process and eliminate the cochannel interference, under given conditions, with minimal computational load, directly from the time-domain received samples, with no DFT/IDFT operations performed prior to the estimation filter. In this way, a large amount of computational load is saved. In the following, a summary of the proposed channel estimator is shown.

The first OFDMA symbol of the transmission packet (preamble) is used to transmit pilots. In our MIMO system,  $N_t \times N_{RS}$  or  $N_{RS} \times N_u$  channels need to be estimated and so, in order to improve the system's efficiency, we propose that the preamble be shared among all transmit paths. From BS or RS, superimposed pilots sequences are sent by the different  $N_t$  transmit antennas (in the case of relays, different  $N_{RS}$  relays). To mitigate the resulting cochannel interference, orthogonal phase-shift sequences are used in each path, where each transmit antenna path uses a distinct pilot sequence  $p_\ell^k$  according to

$$p_\ell^k = \exp\left(-2\pi j \frac{\ell}{N_t} k\right), \quad (14)$$

where  $\ell \in \{0, \dots, N_t - 1\}$  is the index of the BS transmit antenna and  $k \in \{0, \dots, N - 1\}$  is the subcarrier index. For the relay-user link,  $N_t$  in (14) must be replaced by  $N_{RS}$ . Denoting  $r_i(t)$  as the time-domain received signal at relay  $i$  (after removing the cyclic prefix), and considering that in the most common channel models, the taps of the time-domain channel impulse response are uncorrelated and typically limited to a number of nonvanishing terms much lower than the Fast Fourier Transform (FFT) length, since the amplitude of the sequence in (14) is one, at the receiver, the time-domain channel impulse response estimate from transmit antenna  $\ell$  to relay  $i$ ,  $\hat{h}_{\ell,i}$ , is

$$\hat{h}_{\ell,i}(\tau) = r_i(\ell m + \tau), \quad (15)$$

where  $m = N/N_t$  represents the number of samples that are collected from each antenna, and  $\tau \in \{0, \dots, m\}$ . It should be noted that  $m$  is also the limit for the maximum channel delay (normalized to the system's sampling interval). This value is especially important on the second hop, limiting the number of relay channels that can be estimated using only one OFDMA symbol. Going over this limit will result in some performance degradation due to the distortion caused by the cochannel interference. To obtain the frequency-domain channel response, a FFT is applied on  $\hat{h}_\ell$ . Since we use OFDMA, the multiuser channel estimation is performed using only the desired frequencies. This channel estimator

will be denoted throughout the paper as LS, since it follows the LS criterion.

If the channel impulse response estimate contains more samples than the normalized channel length, some of them will only contain noise, and thus these samples will degrade the channel estimation performance. For this reason, we also implement the Most Significant Tap (MST) channel estimation [48], applied to [26], where we only take the most significant  $L$  taps. This low cost improvement of (15) will be denoted as MST throughout the text, and it provides significant performance improvements, especially in the case of LDPC codes, as will be seen in Section 5.

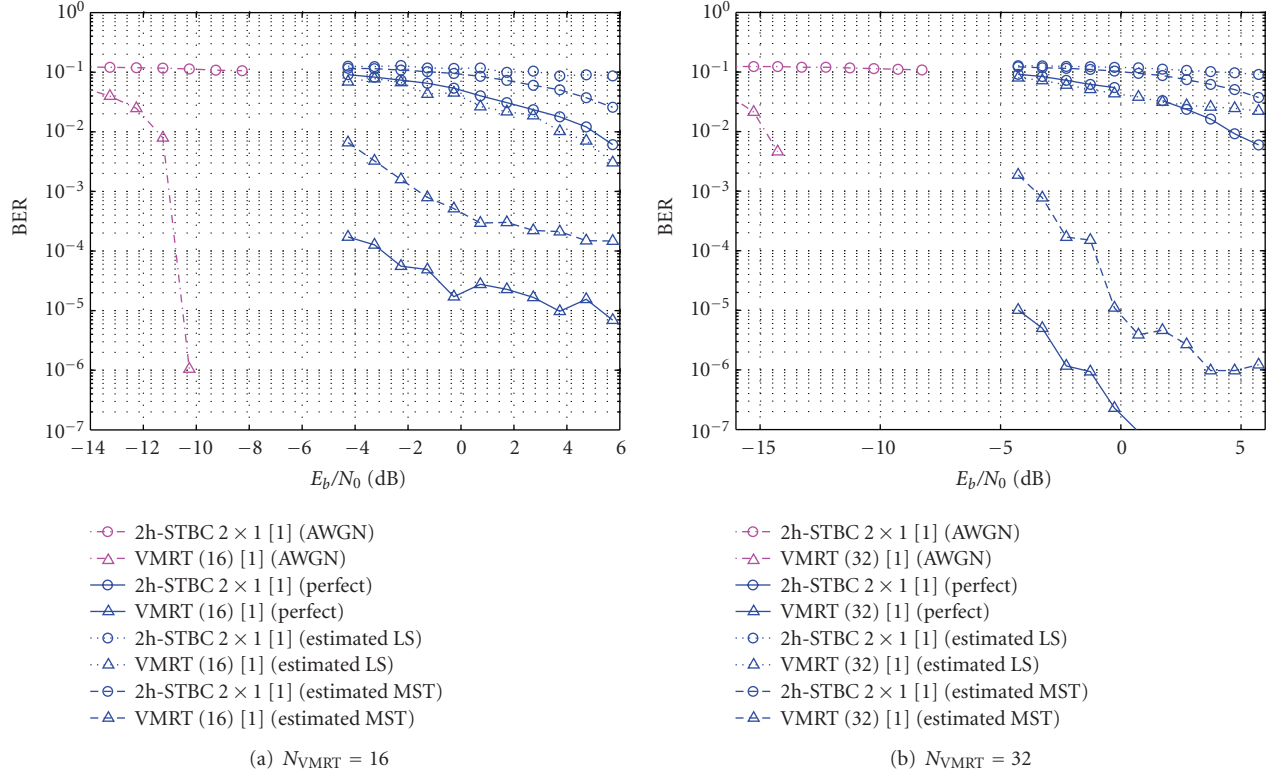
## 5. Simulation Results

Several simulations have been carried out using the Monte Carlo method to evaluate the proposed scheme under realistic channel conditions. All simulations use  $N = 128$  subcarriers and a cyclic prefix of 16 samples over a SUI-3 [51] or HiperLAN 2 B channel model [52]. Since we are not focusing on subcarrier scheduling policies, a block of  $N/N_u$  contiguous subcarriers is assigned to each user. Only user 1 results are presented because similar performance is obtained by the different users, as explained before. In [25], it is shown that we can obtain diversity and array gain on both hops, and this gain increases as the number of RS does. Since this paper is focused on the performance of channel estimation and LDPC codes, we fixed the number of transmit antennas at the BS to 4, the number of relays to 8 (and 8, 16, 32 for LDPC codes), and the number of users to 4. Obtained results can be extrapolated to other configurations because they do not depend on these parameters. The two channel estimation algorithms proposed in the paper, namely, the LS and the MST, have been evaluated in two different scenarios:

- (i) Scenario A: the two links have the same SNR.
- (ii) Scenario B: the SNR of the first link is fixed to 20 and 30 dB for QPSK and 64-QAM, respectively.
- (iii) Scenario C: when using LDPC codes, performance is usually given as a function of the  $E_b/N_0$  (the energy per uncoded bit over the noise). For this reason, results on LDPC will use the  $E_b/N_0$  instead of the SNR. In these cases, the  $E_b/N_0$  for the first link has been fixed to 3 dB.

**5.1. Maximum Ratio Transmission-Single Link (MRT-SL).** Before presenting the results, in the following, a comparison model is introduced. In [7], an optimized transmission scheme based on relays is proposed. The BS uses a single antenna and selects the best relay to transmit to. Then, from this relay the signal is forwarded to the destination. Adapting [7] to be used with multiple antennas at the BS, we have the *Maximum Ratio Transmission-Single Link* (MRT-SL). In this scheme, the BS, based on the channel state information in the link BS-RS, selects the best relay to transmit to and beamforms the transmission to it according to the maximum



FIGURE 5: Effect of channel estimation: BPSK with LDPC codes, SUI-3 channel;  $N_t = 4$ ,  $N_{\text{RS}} = 8$ ,  $N_u = 4$ . Scenario C.

ratio transmission criterion [53]. Thus, transmitted signal can be written as

$$\mathbf{X}^k \big|_{\text{MRT-SL}} = \mathbf{V} \big|_{\text{MRT-SL}} \mathbf{C}^k \big|_{\text{MRT-SL}} \quad (16)$$

with  $\mathbf{C}^k \big|_{\text{MRT-SL}} \in \mathbb{C}^{N_t \times N_s} = \text{diag}\{\mathbf{c}^k\}$ ,  $\mathbf{c}^k$  (a column vector with the  $N_s$  data to be sent in this block on subcarrier  $k$ ), and  $\mathbf{V} \big|_{\text{MRT-SL}} \in \mathbb{C}^{N_t \times N_s}$  being the matrix formed by the repetition of  $N_s$  times vector  $\mathbf{v} \in \mathbb{C}^{N_t \times 1}$ , which are the beamforming weights, again, according to the *minimax* criterion. Thus

$$\mathbf{v} = \frac{\mathbf{h}_{i^*}^{k*}}{\|\mathbf{h}_{i^*}^{k*}\|}, \quad i^* = \arg \min \left\{ \max_k \{ \text{BER}_i^k \} \right\}, \quad k = 1 \cdots N, \\ k^* = \arg \max_k \{ \text{BER}_i^k \}, \quad i = 1 \cdots N_{\text{RS}}. \quad (17)$$

Again,  $N_t = N_s$ . It should be noted that here the search is over the whole subcarrier set because the relays need to receive the signal in the whole bandwidth. In this way, only the  $i$ th relay is able to decode the data. Then, from this relay, data are sent to the users in a single-Input single-output (SISO) link; that is,  $\mathbf{W}$  in (4) is  $w_{j,\ell} = 0$ ,  $\forall j \neq i, \ell \neq i$ , and  $w_{i,i} = 1$ .

This scheme follows [7] but is adapted for a scenario with multiple transmit antennas and without MRC performed at the destination. As will be seen later, this scheme does not exploit diversity on the second hop. Indeed, the best relay from the point of view of BS might not be the best one to reach users. It has the advantage that CSI-T is needed at the BS only for the link BS-RS instead of the whole link CSI-T as in [14]. This scheme will be used for comparison purposes.

**5.2. Effect of the Channel Estimation.** Results have been obtained using the channel estimated by the proposed algorithms at each of the steps in the transmission link. For clarity reasons, in the following, the places and purposes of channel estimation are summarized as follows:

- (i) 2h-STBC—where: at the reception of RS and UT. Reason: for coherent demodulation.
- (ii) MRT-SL—where: at the RS receiver. Reasons: to calculate the beamforming weights and for coherent demodulation. Where: at the reception of UT receiver. Reason: for coherent demodulation.
- (iii) VMRT—where: at the reception of RS. Reason: for coherent demodulation. Where: at the UT receiver. Reasons: to calculate the beamforming weights and for coherent demodulation.

It should be noted that for schemes using MRT, the channel estimation errors will produce a twofold effect: first, the beamforming weights will be corrupted by these errors, and second, the coherent demodulation will also be affected.

In Figure 2, the channel estimation effect on different schemes is shown for a QPSK modulation over a SUI-3 channel and the two scenarios. It can be observed that the VMRT scheme outperforms the others. A diversity gain and an array gain can be observed, due to the multiple transmit elements (relays) on the second hop as stated in [25]. In addition, in Figure 2(a), it can be seen that all the schemes behave similarly when the proposed LS channel estimation is used (around 3 dB of loss in SNR with respect

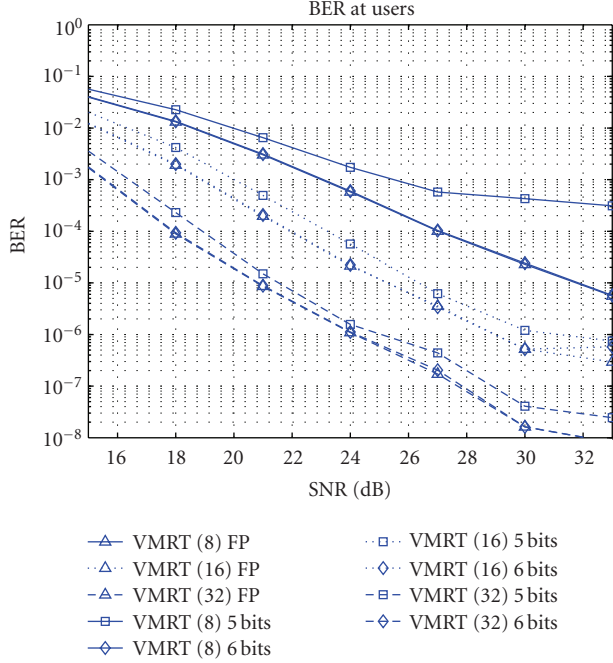


FIGURE 6: Performance results for Uncoded 64QAM. Effect of the quantization on the VMRT.  $N_T = 4$ ,  $N_{RS} = 8$ , and  $N_{VMRT} = 8, 16, 32$ . full precision (FP) and the number of bits for precision.

to a perfect CSI). However, in the case of MST estimation, the gain obtained depends on the scheme and the scenario. In scenario A, by using MST estimation with VMRT, we obtain a gain (with respect to the LS estimation) of around 1.5 dB, whereas for the 2h-STBC, it is around 1 dB, and for the MRT-SL, the gain is less than 0.5 dB. This means that the VMRT scheme is more robust to channel estimation errors, but it is also more sensitive to the algorithm used to estimate the channel. Indeed, the proposed design with MST channel estimation obtains only a degradation of around 1 dB with respect to a perfect CSI. For the results on Scenario B in Figure 2(b), there is a gain of 3 dB for the VMRT, around 2 dB in the case of 2h-STBC and 1 dB for the MRT-SL. Thus, it can be concluded that channel estimation errors affect the coherent demodulation more than the weight calculation. The reason is because for the 2h-STBC (which will only exhibit the coherent demodulation effect), once the SNR in the first link has been fixed to a relatively good value, the MST obtains 0.5 dB of degradation with respect to the perfect CSI knowledge, whereas for the VMRT (which calculates the weights in the second hop), the degradation of MST performance with respect to the perfect CSI is around 0.2 dB. This is mainly due to the coherent demodulation errors in the first link. Furthermore, it can also be observed that there is an error floor caused by the errors on the first link that cannot be recovered, although this error floor is lower (around  $7 \cdot 10^{-8}$ ) for the VMRT than for the other schemes (around  $3 \cdot 10^{-6}$ ).

Similar results are obtained when 64-QAM modulation is used over a SUI-3 channel, as can be observed in Figure 3, which is interesting since results do not depend on the

modulation order; there is only a shift in the SNR values for QPSK with respect to 64-QAM.

Next, in Figure 4(a), the same results as in Figure 3(a) are presented but over an HiperLAN 2 B channel (more frequency selective behavior, used to check the robustness of the scheme and the channel estimator). It can be observed that the estimator is robust and accurate even for a highly frequency-selective channel.

**5.3. LDPC and Channel Estimation.** Recently, capacity-approaching LDPC codes [24] have attracted much attention. Their application to relay-based networks has also recently attracted interest [54–59], although, to the authors' knowledge, the performance has always been evaluated in AWGN scenarios: for single-carrier, single-relay and single-antenna half-duplex transmission in [54, 57], when relays re-encode the signal, and in [58] when they do not, and for multiple-antenna in [55]. If there are many relays conforming a virtual transmitter (although scenarios proposed by those authors only take into account a few), in [56], the increase in performance is noticeable. In [59], the work in [57] is applied to multicarrier signals.

It is well known that random puncturing degrades the LDPC codes performance, and so, in a relay-based system with a realistic channel estimation algorithm, this situation might occur very often. It would be interesting to show how the global performance, when using powerful forward error correction (FEC) such as LDPC codes in the system, would be affected by the channel estimation strategies, and how it does so in the proposed transmission schemes. A similar rate 1/2 LDPC code as in IEEE 802.16e standard [60] is used. As can be seen in Figures 4(b) and 5, several interesting aspects can be found. The first one is that our proposed scheme, combined with LDPC, in AWGN channels, obtains a large gain. The coding gain of LDPC together with our diversity and array gains gives a relay-based system that is able to work with very low  $E_b/N_0$  in both links. The second one is that the scheme still works in wireless channels such as SUI-3, although with an increase in BER and a decrease in diversity gain. The third one is that the channel estimation errors seriously affect the global performance of LDPC codes, and thus it is important to improve channel estimation algorithms to boost performance. Our proposed efficient and simple MST algorithm is able to improve the performance, although there is still a 3 dB penalty with respect to a perfect CSI.

**5.4. Effect of the Feedback Quantization.** Another important aspect is the number of bits needed for the quantization of the weights in the VMRT scheme. In Figure 6, the effect of the number of bits on a fixed point feedback is shown. It can be observed that if the number of bits is too low there is a degradation in the performance (an error floor may even appear), but once the number of bits is sufficient (and not very high), the system performs almost the same as in the case of using full precision. In addition, it can also be appreciated that the degradation decreases with a large number of relays. The reason is because when increasing the

number of relays, quantization errors may compensate one another. It should be noted that, although it is a multiuser MIMO system, on the second hop, only one user feeds the weights back (the selected one), so the feedback does not depend on the number of users but only on the number of transmit elements ( $N_{RS}$ ). This is indeed another advantage of this cooperative scheme.

Moreover, in order to compress the feedback requirements even more, the value of the *quality* of each user can be quantized. It has been shown in [61] that with one or two bits (per user) it is enough to reach more than 95% of the possible throughput.

## 6. Conclusions

In this paper, the scheme denoted as *Virtual Maximum Ratio Transmission* for a cooperative MIMO-OFDMA-relay-based network is evaluated in the presence of realistic propagation channels such as SUI-3 or HiperLAN 2 B channel models. A practical, simple, and efficient multiuser MIMO channel estimation algorithm and the use of LDPC codes are also analyzed.

It has been shown that the scheme is robust against channel estimation errors and it still provides diversity and array gains in such scenarios. Furthermore, when combined with powerful channel codes such as LDPC, the joint advantages result in a significant improvement, allowing the coverage extension even if the first-link SNR is very low. Thus, the VMRT is a cooperative transmission scheme that can increase coverage and system throughput without increasing users' hardware and/or complexity.

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## Preliminary call for papers

The 2011 European Signal Processing Conference (EUSIPCO-2011) is the nineteenth in a series of conferences promoted by the European Association for Signal Processing (EURASIP, [www.eurasip.org](http://www.eurasip.org)). This year edition will take place in Barcelona, capital city of Catalonia (Spain), and will be jointly organized by the Centre Tecnològic de Telecomunicacions de Catalunya (CTTC) and the Universitat Politècnica de Catalunya (UPC).

EUSIPCO-2011 will focus on key aspects of signal processing theory and applications as listed below. Acceptance of submissions will be based on quality, relevance and originality. Accepted papers will be published in the EUSIPCO proceedings and presented during the conference. Paper submissions, proposals for tutorials and proposals for special sessions are invited in, but not limited to, the following areas of interest.

## Areas of Interest

- Audio and electro-acoustics.
- Design, implementation, and applications of signal processing systems.
- Multimedia signal processing and coding.
- Image and multidimensional signal processing.
- Signal detection and estimation.
- Sensor array and multi-channel signal processing.
- Sensor fusion in networked systems.
- Signal processing for communications.
- Medical imaging and image analysis.
- Non-stationary, non-linear and non-Gaussian signal processing.

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Proposals for special sessions	15 Dec 2010
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